Communication by Means of Reflected Power*

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Summary—**Point-to-point communication, with the carrier power generated at the receiving end and the transmitter replaced by a modulated reflector, represents a transmission system which possesses new and different characteristics. Radio, light, or sound waves (essentially microwaves, infrared, and ultrasonic waves) may be used for the transmission under approximate conditions of specular reflection. The basic theory for reflected power communication is discussed with reference to conventional radar transmission, and the law of propagation is derived and compared with the propagation law for radar. A few different methods for the modulation of reflectors are described, and various laboratory and field test results discussed. A few of the civilian applications of the principle are reviewed. It is believed that the reflected-power communication method may yield one or more of the following characteristics: high directivity, automatic pin-pointing in spite of atmospheric bending, elimination of interference fading, simple voice-transmitter design without tubes and circuits and power supplies, increased security, and simplified means for identification and navigation.**

I. Radar Transmission with Scattering Target

IN THE CONVENTIONAL radar application, the return radiation from the target carries the information that the target exists. In the simplest case, therefore, the radar receiver response indicates a "yes" or "no," and the type of modulation employed may be considered as being of the "on-off" type. The following paper concerns utilization of reradiation from a target when the target is subjected to any kind of modulation; in particular, voice and telemetering-data modulation.

The geometrical configuration, size, shape, and surface conditions of the target determine to a considerable extent the law of propagation for the chosen type of transmission. In the conventional case of radar transmission with scattering target, the propagation follows basically an inverse-fourth-power law, which may be written:

$$d_{max} = \sqrt[4]{\frac{A^2k^2}{4\pi\lambda^2}\sigma\frac{P_T}{(P_R)_{min}}} \tag{1}$$

where

d = distance from transmitter-receiver to target
A = aperture of transmitting antenna
k = dimensionless factor, depending upon efficiency of antenna aperture
λ = wavelength of transmission
σ = radar cross section of target[1]
P_T = transmitted pulse power
P_R = received power.

The minimum received power $(P_R)_{min}$ which will give satisfactory radar operation over a maximum distance d_{max} is determined by a number of factors, some of which will be discussed later. For simplicity, the factors under the radical sign A, k, and λ may be assumed constants. It then follows that the range of the radar depends on the radar cross section of the target, and the ratio of transmitted pulse power to minimum received power, required for satisfactory radar operation. The result may not indicate the ultimate value of d_{max} for the reason that (1) is merely the prediction on paper of a relationship between transmission characteristics. This relationship is rather complicated in practice.

The signal-to-noise ratio in a conventional radar with A-scope presentation is boosted by integration performed by the human eye, although other integration may be utilized. It is of interest to study the maximum distance that can be obtained with fourth-power propagation, assuming a reasonable time T_{tot} available for integration, such as $T_{tot}=1$ minute. This maximum distance can then be compared with that achieved with second-power or better propagation, obtainable in communication using a properly modulated nonscattering target. Consider first the signal-to-noise ratio when no integration is present. The random noise power (thermal noise, shot-effect noise) is proportional to the bandwidth B or inversely proportional to the pulse duration τ_P, so that for a noise amplitude N

$$N \sim \sqrt{B} \sim \sqrt{1/\tau_p}. \tag{2}$$

The transmitted energy is proportional to $h^2\tau_p$ when h is the height of the pulse, and also proportional to $P_{Tav}T_0$ where P_{Tav} is the average transmitted power and T_0 is $1/f_{rep}$. Thus

$$h \sim \sqrt{\frac{P_{Tav}T_0}{\tau_p}}. \tag{3}$$

With h indicating the signal S, the ratio of (3) and (2) will give the signal-to-noise ratio

$$\frac{S}{N} \sim \sqrt{P_{Tav}T_0}. \tag{4}$$

This expression is independent of the pulse width or the corresponding bandwidth.[2] What this means for constant P_{Tav} is that the pulse height varies accordingly; the loss in height of the pulse being compensated for by the increase in pulse length. The signal-to-noise ratio is not independent of the $prf(f_{rep})$, however, for a reduced prf will give a larger T_0 in (4), thus an increased signal-

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[1] The radar cross section of the target is defined as follows:

$$\sigma = 4\pi\frac{\text{power per unit solid angle scattered back towards the transmitter}}{\text{power per unit area in wave incident on target}}.$$

[2] S. Goldman, "Some fundamental relations concerning noise reduction and range in radar and communication," PROC. I.R.E., vol. 36, pp. 584–595; May, 1948.

to-noise ratio. In practice, there is a definite limit to T_0 and (4) then gives the corresponding limit in signal-to-noise ratio.

It should be noted that, due to signal suppression by noise in the receiver detector ("second detector"), i.e., because of lost coherence,

$$\left(\frac{S}{N}\right)_{post\text{-}rect.} = \frac{\overline{S}}{\overline{N}} \sim \left(\frac{S}{N}\right)^2_{pre\text{-}rect.} \tag{5}$$

where the bars indicate postrectification or output quantities.

Consider next the case when integration is used in connection with a gate that only opens up for the duration of the pulse. The signal-to-noise ratio then becomes insignificant, and the quantity of interest becomes the ratio between the signal $\overline{S}$ and the noise fluctuation $\overline{\Delta N}$ in the integrated output. It is known in probability calculus that the square of the fluctuation is proportional to the stochastic quantity, which in this case is the number of integrated pulses T_{tot}/T_0, while the mean itself is proportional to the stochastic quantity, thus

$$\overline{\Delta N} \sim \sqrt{\frac{T_{tot}}{T_0}},$$

$$N \sim \frac{T_{tot}}{T_0},$$

and, therefore,

$$\frac{\overline{\Delta N}}{\overline{N}} \sim \sqrt{\frac{T_0}{T_{tot}}}. \tag{6}$$

It follows from (4), (5), and (6) that

$$\frac{\overline{S}}{\overline{\Delta N}} \sim \frac{(S/N)^2}{\overline{\Delta N}/\overline{N}} \sim \sqrt{P_{Tav}T_0}\sqrt{P_{Tav}T_{tot}}. \tag{7}$$

Theoretically, we could see any target if it could be studied during a sufficiently long time—an hour, or a day. In (1), this corresponds to reliable radar operation with a greatly reduced value of $(P_R)_{min}$. The technique applied is the technique of integration, or signal storage. While a follow-up on this matter would lead us outside of the scope of this paper, a few points of interest should be mentioned. The result in (7) indicates that the maximum distance depends upon the total transmitted signal energy $P_{Tav}T_{tot}$, which controls the response of the integrator, and also indicates that the signal-to-noise-fluctuation ratio increases with the period of pulse repetition for given average transmitter power, which should be as high as possible. With recently developed storage methods, the fixed-target range may be increased as much as four times,[3] which is significant in the comparison of inverse-fourth-power-law radar transmission with more favorable communication transmission, as storage methods at present scarcely apply to the communications field. Under practical operating conditions, with the best possible integration system, the inverse-fourth-power law nevertheless restricts the range severely, and it is of the greatest interest to study the conditions under which a more favorable propagation law can be obtained.

[3] F. Dickey, T. A. G. Emslie, and H. Stockman, "Storage of Signals in Noise," unclassified report, in preparation, USAF, AMC, ERL, Cambridge, Mass.

II. Communication Transmission with Nonscattering Target

Consider the conventional communication system in Fig. 1(a), in which the signal modulates the radiation from a transmitter at B and the intelligence is recovered from the radiation at A. This arrangement may be compared with the new system, Fig. 1(b), in which the source of radiation is located at the point of reception A, and the carrier power is reflected back from B by means of a signal-excited reflector; modulation taking place at the point of reflection B. If two-way communication is desired, the radiation source at A may be modulated, or all equipment duplicated.

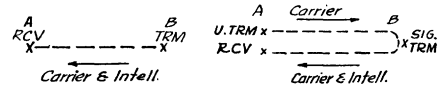

Fig. 1—A comparison between conventional direct-power communication (a) and the new reflected-power communication (b), in which reradiation provides the carrier for the signal.

It is of interest to study the relation in (1) in a more simple and general way that is not particularly restricted to the fourth-power law, and it is believed that the following expression describes the conditions of immediate interest:

$$\frac{P_R}{P_T} = k_1 f(\lambda, d) k_r \frac{1}{d^n} \tag{8}$$

where k_1 is a proportionality factor, k_r a factor describing the reflecting characteristics of the target, and n an exponent, which in the case of conventional radar transmission has the approximate value 4. The quantity $f(\lambda, d)$ indicates the reduction in P_R relative to P_T due to the chosen values of wavelength λ and distance d, but for the time being the only variation in P_R considered is that due to the spreading of the beam. While (1) referred to radio waves, (8) is general and applies to any kind of transmission, and particularly "light"-wave (infrared), and sound-wave (ultrasonic), transmission. The unmodulated transmitter may be a magnetron, klystron, infrared lamp, or ultrasonic whistle. It is now required that the wavelength λ of the chosen transmission and the equivalent area A_e of the target (in the form of a modulated reflector) fulfill the requirements

$$\lambda^2 \ll A_e, \tag{9}$$

$$\alpha_{view} \gg \alpha_{diffr} \sim \frac{\lambda}{\sqrt{A_e}} \tag{10}$$

where α_{view} is the angle through which the source of radiation sees the target, and α_{diffr} the diffraction beam width corresponding to the equivalent reflector area A_e.

While these conditions may be very difficult to fulfill even for a K-band radar (λ of the order of 1 cm), they may be easily fulfilled for "light"-wave transmission. The geometrical relationships are then the ones shown in Fig. 2, where x indicates the position of the source of radiation, y the position of the reflector with area $A_e = S$.

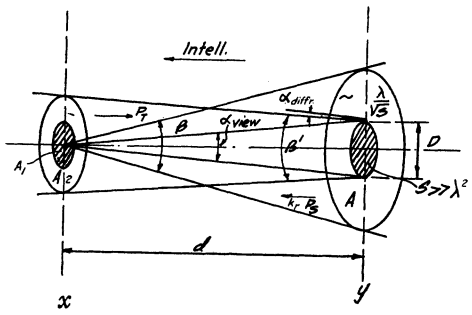

Fig. 2—The geometry of specular reflection, showing that for equal size antennas the reduction in power during the return path is only in a ratio of 1:4.

If P_T is the power radiated within the beam described by the angle β, and if uniform power distribution is assumed at y, then for $k_r = 1$ the reflected power P_S becomes

$$P_S = \frac{S}{A} P_T = k \frac{S}{d^2} P_T \tag{11}$$

where k depends on β only ($k = 1/\pi \tan^2 \beta/2$). This is essentially a derivation applicable to both the radar and the communication case, and proves the well-known square-law relation for one-way propagation. In the radar case we have, in general, $\alpha_{diffr} \gg \alpha_{view}$. In that case the reflector acts as a scatterer and the transmission in the direction yx is similar to the direct radiation, being characterized by a fixed beam angle β'. Thus, repeating the procedure xy in the direction yx, we obtain

$$\left.\frac{P_R}{P_T}\right|_{radar} \sim \frac{1}{d^4}. \tag{12}$$

Repeating the procedure in the communication case, we reradiate a beam described by the angle α_{view} and accordingly obtain an illuminated area $A_2 = 4S$. Thus

$$P_R = \frac{A_1}{A_2} P_S = \frac{A_1}{4S} k \frac{S}{d^2} P_T, \tag{13}$$

or, for $A_1 = S$,

$$P_R = \frac{1}{4} k \frac{S}{d^2} P_T.$$

This means that, while the loss in power in the radar case is very great in the direction yx, perhaps of the order of 10^6, *it amounts only to a factor of four* for very short-wave communication in the same direction, i.e., in the direction the signal is being carried. In contrast to (12), the condition for communication is

$$\left.\frac{P_R}{P_T}\right|_{comm} \sim \frac{1}{d^2}. \tag{14}$$

If the target picks up all transmitted radiation and if the receiver antenna is also built to pick up all the transmission, i.e., if the receiving antenna (for specular reflection) has twice the diameter of the target, there will be no loss due to the spreading of the beam, so that $n = 0$ in (8). Practical signaling has actually been carried out over a test-range distance of 150 yards[4] with $n = 0$. Square-law signaling is the best one can hope for in most practical applications, but the difference between square-law signaling and fourth-power signaling is very considerable. If the fourth-power law yields useful transmission over 10 miles, square-law transmission will increase the range to 100 miles, on the assumption that everything else remains equal. With an inverse-square-law propagation, increase of the transmitted power is a practical consideration, as only 4 times more power is needed for doubling of the range compared to 16 times for the inverse-fourth-power law.

Further comparison of inverse-square-law propagation (using light waves, $\lambda = 5 \times 10^{-5}$ cm) and inverse-fourth-power-law propagation (using microwaves, $\lambda = 1$ cm) is made in Fig. 3.[5] The transmitted beam width is 2°, due to diffraction in the microwave case, and to finite size of the source in the optical case. The light waves show an advantage in received power of 74 db over the microwave transmission at a distance of 90 km. This figure, of course, does not include the relative efficiencies of transmitters and receivers in the two cases. Further comparison between light-wave and microwave transmitters and receivers must be made before practical conclusions can be drawn.

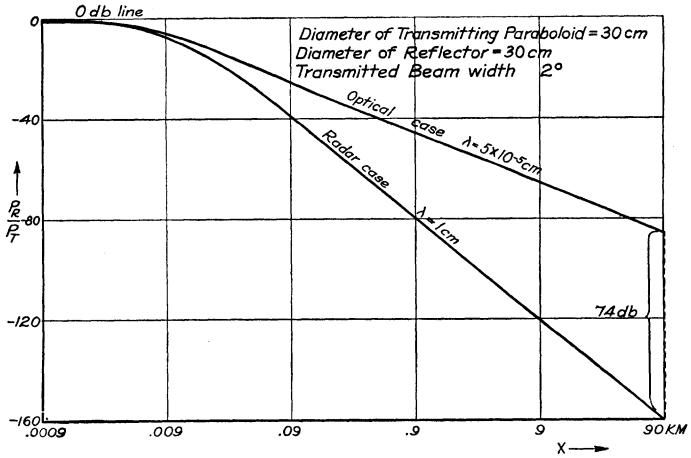

Fig. 3—A comparison of inverse-square-law and inverse-fourth-power-law propagation, showing an advantage of 74 db of the former over the latter at 90 km distance under assumed conditions.

It is of particular interest to study the increase in range of beacon operation over conventional radar operation. In the general beacon application the target is a receiving antenna, which picks up the signal and uses it to trigger a beacon transmitter, which reradiates a coded signal to the radar transmitter. The transmissions

[4] See Part IV, Measurement Results, of this paper.
[5] From values calculated by A. G. Emslie.

are therefore of square-law nature in both directions. We may write for the interrogating direction of transmission, using the index 1 for the radar that serves as the source of transmission, and index 2 for the beacon that serves as the receiver,

$$\frac{(P_R)_2}{(P_T)_1} = \frac{G_1G_2\lambda^2}{16\pi^2}\frac{1}{d^2} \tag{15}$$

where G_1 and G_2 are the gains of the radar and beacon antennas, respectively.[6] Similarly, in the other direction

$$\frac{(P_R)_1}{(P_T)_2} = \frac{G_1G_2\lambda^2}{16\pi^2}\frac{1}{d^2}, \tag{16}$$

so that, independent of distance,

$$\frac{(P_R)_2}{(P_T)_1} = \frac{(P_R)_1}{(P_T)_2}. \tag{17}$$

Here $(P_R)_1$ must be sufficiently large for reception. For a given $(P_T)_1$ the received power in the beacon $(P_R)_2$ is determined by the inverse-square-law propagation. If $(P_T)_2$ is assumed equal to $(P_T)_1$, then $(P_R)_1=(P_R)_2$, and no extra power loss is encountered due to the round trip. If $(P_T)_2$ is made just large enough to provide sufficient value of $(P_R)_1$, called $(P_R)_1'$, the replacement of the conventional beacon system with an ideal reflected-power system reduces the received power only in the ratio of 1:4, or to $0.25(P_R)_1'$, which is not a serious loss. This statement does not infer that a reflected-power beacon system is better than, or compares favorably with, a conventional beacon system, but serves to direct the attention to the fact that square-law propagation is obtainable with both systems. It should be noticed that, while the conventional beacon system replies in all directions, the ideal reflected-power beacon system only replies in the direction of interrogation. Whether or not a suitable reflected-power beacon system can be built in practice remains to be seen.

III. Methods of Modulation

The target or reflector may be modulated in a number of ways, of which only a few will be mentioned: variable-damping modulation, interference or phase modulation, directional modulation, position modulation, doppler modulation, and polarization modulation. The first three will be discussed in the text.

The methods of modulation concerned apply particularly to corner reflectors. A corner reflector has the important property that a ray, which enters the corner, will experience a reflection from each of the surfaces, and will return in the direction from which it came.

As seen from the source of radiation, the corner reflectors[7] will show regions of single, double, and triple reflection. The triple-reflection region always provides a return radiation coincident with the incident radiation and is divided into six sections in circular arrangement. This radiation has a plane wave front, the corner reflector serving effectively as a plane mirror. The triple reflection has maximum intensity in the direction which makes equal angles with each axis of the co-ordinate axis system described by the edges of adjacent reflecting walls. Radiation entering at an angle deviating from the optimum angle causes less reflection, and the intensity of the reflected radiation tapers off gradually until triple reflection reaches zero when the line of sight lies in a reflecting plane. Some of these reflector characteristics, particularly those of interest for modulation, have been investigated in laboratory measurements and field tests to be described later.

Some of the difficulties encountered in reflector modulation with particular reference to corner reflectors are as follows. The reflector must be large to yield sufficient power return, and the required size increases with the wavelength (see the condition (9)). Modulation usually requires mechanical oscillation of large masses, joined into a rigid system by members of insufficient stiffness; thus the upper modulation cutoff frequency becomes unduly low. In addition, each reflector yields a particular receiver response curve, i.e., Fourier spectrum, for the applied modulation. Efforts to improve the radiation pattern may conflict with efforts to improve the Fourier spectrum. Thus arrangements to provide omnidirectional response may conflict with the requirement of signal response on the fundamental only. Conditions are complicated by additional requirements; for example, the requirement that stray radiation must not exceed a certain db level, etc.

It is desirable that new types of reflectors be made available which can be modulated by video signals up to cutoff frequencies of the order of 5 or 10 Mc. While such reflectors are not available today, methods of design are beginning to appear, which, for light-wave transmission in particular, may provide modulation response for frequencies far above the audio range. The interference type of modulation may be used advantageously, but the investigator must be cautioned by the fact that the incoherent nature of light restricts the freedom of choice of amplitudes and reflector characteristics in a modulator of this kind.

Variable-damping modulation may act upon both the impinging and the reradiated wave. A reflector arranged for variable damping may be looked upon as a parasitic equivalent circuit with variable admittance, so that the response E_r from the reradiated field may be of the form (18), where E_i represents the impinging field, k is the reflection coefficient in absence of modulation, and a, b, $\cdots$, coefficients indicating the nature of the frequency response to the modulation frequency $\Omega/2\pi$.

$$E_r = kE_i(1+a\cos\Omega t+b\cos 2\Omega t+\cdots)\cos(\omega t+f_m(t)). \tag{18}$$

The formula shows that we may expect not only a somewhat distorted amplitude modulation, but also phase modulation $f_m(t)$.

[6] L. N. Ridenour, "Radar System Engineering," MIT Rad. Lab. Series, McGraw-Hill Book Co., New York, N. Y., 1947.

[7] See brief discussion of corner reflectors, Chapter 3.5 of footnote reference 6.

Interference or phase modulation may be produced by varying the distance between two reflecting surfaces; for example, by oscillating one of the surfaces by means of the sound waves produced by the human voice. The vectors representing the reradiations from the two surfaces will have a variable phase difference controlled by the modulation. There are regions of linear modulation as well as distorted modulation. As an example, for the case of a 90° separation angle, changed by a certain amount ϕ by modulation, the two vectors $\bar{a}_1$ and $\bar{a}_2$ may be assumed to have equal amplitudes A and may be written

$$\bar{a}_1 = A\epsilon^{j\omega t}, \tag{19}$$

$$\bar{a}_2 = A\epsilon^{j(\omega t+\pi/2+\phi \cos \Omega t)} \tag{20}$$

where $\Omega/2\pi$ is the modulation frequency. The resulting vector magnitude then becomes

$$|\bar{a}_1 + \bar{a}_2| = A\sqrt{2(1+\sin\phi)}. \tag{21}$$

For a 30° deviation, permissible in low-fidelity systems, the modulation percentage becomes 20 per cent.

In the case of light-wave transmission the distance between the two surfaces must not be so large that the coherence is lost, in which case reliable interference modulation becomes impossible.

Directional modulation is provided if a beam component is made to describe an angular displacement, which is controlled with respect to amplitude, frequency, and wave form by the modulating signal, so that a "hit and miss" action is provided at the point of reception (see Fig. 4). Here the modulated target, or reflector, is shown to the left with the angle of displacement θ. The response

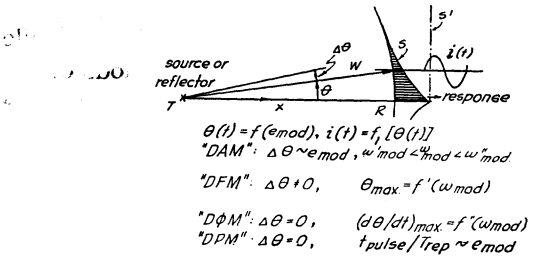

Fig. 4—An example of directional modulation, in which a beam component is made to deviate in accordance with directional amplitude (DAM), frequency (DFM), phase (DϕM), or pulse (DPM) modulation.

with respect to $\theta(t) = f(e_{mod})$ is, as an example, the one shown to the right; so that, in the simplest case, the more we deviate, the weaker becomes the response $i(t)$. Different kinds of transmissions can be established, such as directional amplitude modulation, "DAM"; directional frequency modulation, "DFM"; directional phase modulation, "DϕM"; and directional pulse modulation, "DPM." The differences between these transmissions is indicated by the formulas shown in Fig. 4, but will not be further discussed in this paper.

With the receiving equipment to be described later, it is possible to detect as small a modulation percentage as 0.1 per cent and still maintain a reasonably reliable signal from a modulated reflector. Some of the above modulation methods yield modulation percentages in excess of 10 per cent, and where noticeable distortion is tolerable, in excess of 20 or 30 per cent. Thus, conditions for reliable reception of modulated reflector signals seem to be at hand.

IV. Measurement Results

Practical microwave measurements at the Ipswich Antenna Station in Massachusetts indicate that low-frequency identification modulation (produced, for example, by slowly repeated deviations of the corner reflector around a fixed axis) sometimes must be made as large as 20 per cent to become distinguishable to the receiver operator against the background noise (or "grass") on a radar A-scope. The difficulty in producing a strong and suitable reradiation at ranges of the order of 10 to 50 miles lies in the fact that good propagation characteristics are hard to obtain for radio waves with a wavelength of the order of 1 cm or less, and already at these values (9) and (10) will require corner reflectors of large dimensions. Light waves, such as infrared waves, would meet the requirements of (9) and (10) with small-size reflectors of high modulation cutoff frequency, but in the near-earth atmosphere light-wave transmission is generally practical only for short ranges, of the order of 10 miles or so. Here the possibilities described with reference to (8) with $n=0$ become of particular interest.

Fig. 5 shows a block diagram of a measurement setup suitable for microwave field tests on modulated reflectors, and Fig. 6 shows a photograph of associated equipment.[8] A conventional receiver for the proper transmission frequency and bandwidth is used, followed by a video amplifier and gating amplifier. The gate is obtained from a gate generator, synchronized with the transmitter. When so required, a box generator can be included in the circuit and is shown in the form of a holding circuit, generating the wave form shown. The amplifying system is provided with automatic gain control, and the controlled amplifier output feeds into one

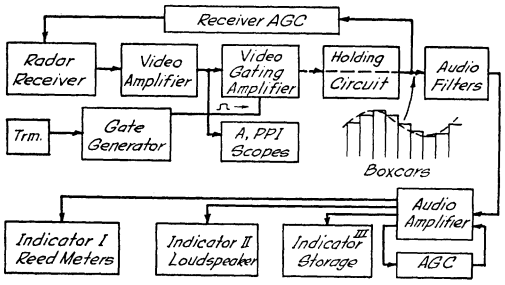

Fig. 5—A simple block diagram of a microwave measuring setup for the investigation of different kinds of reflectors, essentially for K-band operation.

[8] For further information on measurement results and applications, see forthcoming Electronics Research Laboratory report by H. Stockman, entitled "Reflected Power Communications."

or more audio filters, one for each audio channel, connected to an individual audio amplifier. Suitable cathode-ray oscilloscopes are provided for the study of the

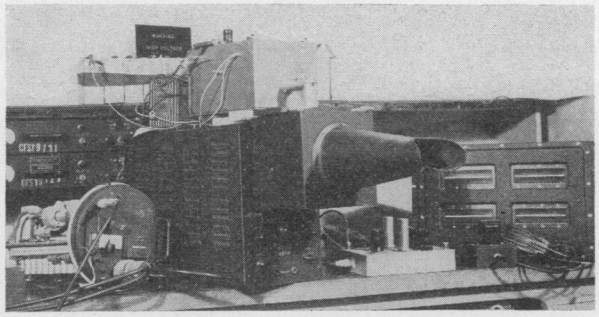

Fig. 6—A photograph of the measuring setup in Fig. 5, showing to the right a Frahm frequency meter used for identification of reflectors modulated with a complex wave. (The NIT system of identification.)

video outputs. Each audio amplifier has a separate automatic gain control. Three different kinds of indicators are shown, marked 1, 2, and 3. The first indicator is a reed meter, the second is a loudspeaker, and the third is a storage device, particularly the mechanical integrator,[8] suggested by the writer for extraction of very weak signals from noise.

The use of this and similar receiving systems has indicated particular requirements on certain receiver characteristics. High receiver gain is necessary, as a receiver for modulated reflector signals should give full output for a very weak incoming wave, modulated less than 1 per cent. Suitable manual and automatic gain controls must be provided, as saturation may remove small amounts of modulation, and the time constants of the control systems must be made very small. Gain control is preferably established by means of a degenerative system. Suitable detection is an important requirement, and rectification, wherever used, should be under full control and of correct form. Linearity in amplifying parts and in the envelope region of rectifying parts must be maintained so that cross-modulation effects are avoided. Optimum gate width must be obtained, and adjustable gate width is preferable, as for very weak signals a somewhat wider gate may give better results than a narrower gate, the opposite being true for ordinary signals. Generally, a wide gate is undesirable from the viewpoint of interference. Freedom from extraneous frequencies is an important requirement, and the receiver must be designed not to yield distortion that produces such frequencies. Filtering in the output part of the receiving system may be required. Freedom from hum and noise is also of the greatest importance, as even a weak ripple component may be very large in comparison with the detected signal component.

For microwave transmission, the following measurement results are of interest. With a large corner reflector as a target (see Fig. 7), the distance of transmission was reduced until a saturation target was obtained with just noticeable "grass" on the base line of an A-scope (distance appreciably one mile). When one of the walls was deviated 5° from the proper right-angle positions, the saturation target almost disappeared from the scope,

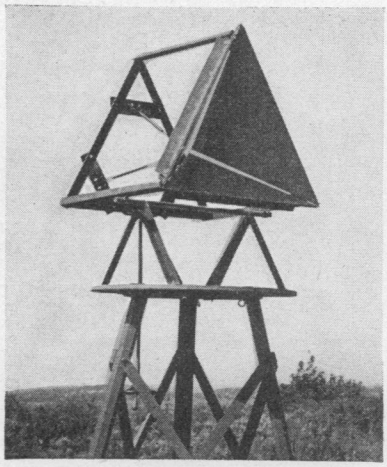

Fig. 7—One of the largest corner reflectors used during the tests with a 2-foot free edge. The sides can be deviated manually or by means of a motor-driven device and readings taken of the response in the receiver.

corresponding to an output modulation of nearly 100 per cent. This result varied with the position the reflector held with respect to the average ground clutter. Very much smaller deviations were found satisfactory on various types of reflectors for good A-scope detection of the target and reliable observation of the code signal. Most of the experiments were carried out with an S- or X-band radar, so as to provide unfavorable predictions from (9) and (10) in order to study the limitations of the method.

With the reed meter and simultaneous reception from several coded reflectors, visible deviations of several reeds on the frequency-meter indicator were observed, superimposed upon the random variations caused by noise.

The following measurements concern primarily the use of modulated reflectors for the purpose of identification by means of reed indicators and refer to a system for reradiation of three or four digits upon excitation by coherent or incoherent electromagnetic radiation.[9] Each reflector is identified by its "numberplate" code number, the digits of which in the first experiments have been chosen in the frequency interval 10 to 100 cps. The measurement results indicate a required minimum frequency spacing of 2 cps, so that, as an example, a suitable code number would be 17, 19, 25. The choice of numbers is limited by the fact that the positions of the digits in the number are without meaning; thus, 19, 17, 25 being identical with 17, 25, 19, and 25, 19, 17, etc. Further, due to the unavoidable nonlinearity in the signaling system, harmonic, sum, and difference tones are produced in the receiver output, and may give false signals. For a

[9] The NIT "Number Identification Target" system suggested by the writer. See forthcoming Electronics Research Laboratory report.

number of reed frequencies $n=90$ and a number of digits $m=3$, the number of combinations or possible code signals n_c may be estimated as

$$n_c = \frac{n!}{m!(n-m)!} = \frac{90!}{3!(90-3)!} \approx 100{,}000. \quad (22)$$

Because of limitations, such as those mentioned above, the practical value of n_c would probably be reduced to the order of 10,000, or so.

Various experiments with "numberplate" coded reflectors have been carried out, and Fig. 8 shows one of the early models. Three turrets of four corner reflectors in each turret are arranged coaxially and rotate with different speeds, driven by the same motor. Both amplitude- and phase-modulation type turrets have been used, and the measurements indicate that phase-modulated turrets provide the most consistent and steady echo returns. Various motor speeds were used, and a Frahm frequency meter with a frequency range from 15 to 85 cps was connected to the receiver (see Figs. 5 and 6). An X-band (3-cm wavelength) radar type AN/APS-3 was used as the radiation source and the distance of transmission varied within a maximum range of 2 miles. The amounts of harmonic generation and transmission radiation pattern were studied, but sufficient results have not yet been obtained. It is indicated, however, that although the experimental turret reflector in Fig. 8 produced a weak second harmonic and a still weaker third harmonic, it fulfilled the purpose of yielding an indication in the Frahm frequency meter from which the "code" number of the reflector could be read off without risk of false indication. The possibility was investigated of surrounding the triple-turret reflector with a dome of insulating material of such thickness that a filter action resulted in a pass band for a particular frequency region. Thus, the reflector would respond only to a particular interrogating beam of radiation.

Fig. 8—Triple-turret reflector, in which each turret rotates with a predetermined speed. The result in the receiver is that particular reeds in the frequency meter become excited.

While the triple-turret reflector provides one solution to the problem, it is possible to generate all three digit frequencies in one and the same reflector, if this reflector is excited by a complex wave composed of the individual waves, or if different surfaces (walls) in the reflector are excited by the individual digit modulation frequencies. Various experiments were carried out and indicated that the suggested principle for complex-wave excitation is useful, but a considerable amount of research and development work remains to be done before practical field tests can be initiated. This investigation should be extended to carrier operation of corner reflectors, the signal modulation being applied to the carrier, and the signal-modulated carrier applied to the reflector.

The possibility of detecting corner-reflector return radiation must be considered with reference to possible existence of ground and troposphere reflections. Destructive and constructive interference is created and depends upon the characteristics and the position of the source of radiation, the receiver, and the reflector. The seriousness of such interference is determined by the operating conditions and the wavelength, and is operationally less severe for a higher frequency than for a lower frequency in the range where reflected power may be utilized.

The most interesting group of measurements concerns small corner reflectors utilizing light-wave transmission, and Fig. 9 shows one of the measurement setups used for a distance of the order of 100 yards. The transmitter to the right has the form of an automobile head light, projecting a narrow beam of light into space via a special periscope and optical lens system. The receiver is located at the same terminal and has essentially the form of a photocell with amplifier, feeding a loudspeaker. The modulated corner reflector to the left is excited by the human voice via a microphone and an amplifier, and the voice current oscillates all three walls in simultaneous action. Measurements with this and other transmission links were carried out and the results indicate that voice communication with good sound quality was feasible, utilizing the reflected-power principle. In later measurements the photocell was turned around and provided with a parabolic mirror and conditions for square-law

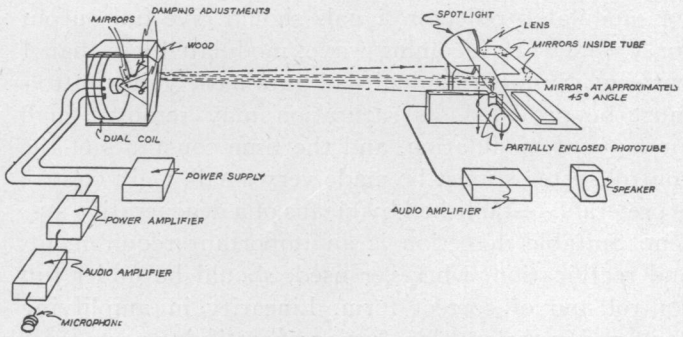

Fig. 9—Optical measurement setup for the study of different kinds of voice-modulated reflectors (to the left), either excited through microphone and amplifier, or directly by the human voice.

propagation investigated. Tests were carried out with the narrow light beam utilized fully in both the modulated reflector and the receiver parabolic mirror, and it was realized that practically all attenuation due to beam spreading could be eliminated. Thus the signal strength was considerably larger than in the square-law case, with $n=2$ in (8), and changes in the distance did not seem to affect the signal strength, as is predicted by (8) with $n=0$.

It was concluded from a theoretical investigation that the power in the human voice would be sufficient to operate a reflector, so that the microphone and amplifier in Fig. 9 could be eliminated. It was indicated that a maximum displacement of 0.01° of the corner reflection wall with an edge length of a few inches would yield a radiated power in the form of air-pressure variations of 5×10^{-6} watts at 400 cps. It is estimated that the average power in conversational speech has a peak value of 5×10^{-3} watts. Comparing the figures 5×10^{-6} watts and 5×10^{-3} watts and allowing a loudspeaker-action efficiency of 10 per cent, it seemed that a reasonable guarantee for direct voice operation of corner reflectors was at hand, assuming that most of the speech power be directed so as to vibrate the corner reflector walls. This could be done by means of a horn operating under proper matching conditions. In view of these predictions a reflected-power transmitter was built for direct voice excitation of all three walls. Later on, additional reflectors were built with one wall replaced by part of a thin dural metal sheet stretched over a frame (see Fig. 10). These reflectors had frequency responses extending from approximately 200 to 4000 cps, somewhat peaked in the region of 3000 cps. The results with these voice-operated reflectors were superior to those obtained with amplifier-operated reflectors, and reliable communication was obtained over the 100-yard test range with good sound quality.

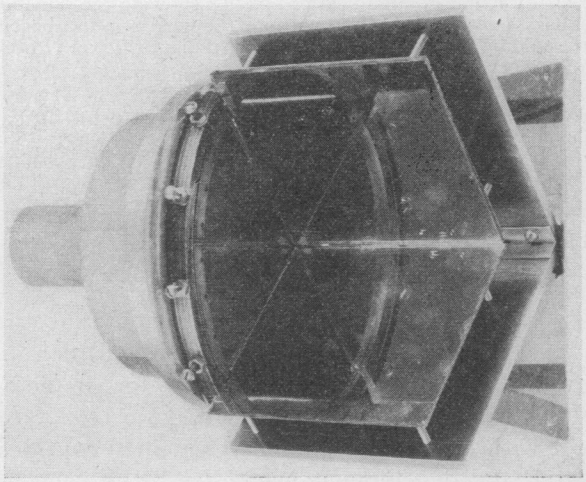

Fig. 10—One of the voice-powered reflectors used during the measurements. Two sides of the corner reflector are glass mirrors, while the third side is a metal diaphragm, excited by the human voice (size of each mirror, 4×4 inches).

V. Practical Applications

The following observations may be made. The source of radiation is basically unmodulated, which invites reconsideration of known, powerful radiation sources, which cannot be easily and properly modulated. Even if the carrier power (Fig. 1 or Fig. 11) is radiated omnidirectionally, the returning modulated radiation may be made highly directive and pin-pointed on the receiver.

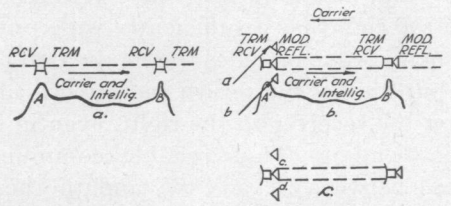

Fig. 11—Conventional microwave relay with direct power (a) and with reflected power (b) and (c). In the latter case the signal-modulated beam is automatically pin-pointed to the receiver even in case of atmospheric bending.

In case the impinging beam or beam component encountered at B in Fig. 1(b) is bent due to atmospheric conditions, the returning modulated radiation remains nevertheless pin-pointed on the receiver, as this radiation utilizes the same path of transmission. The reflector at B does not radiate unless excited by an impinging wave, and it basically operates linearly, in accordance with the superposition theorem, thus yielding freedom from overloading and cross modulation. As the reflector at B may be excited directly by sound waves in the air, the new system makes possible the design of small voice transmitters not using tubes, circuits, or power supplies. Due to the fact that the transmitter and receiver at A are located side by side, various forms of control circuits can be introduced between the transmitter and receiver for one purpose or another, such as secret communication, improved signal-to-noise ratio, and reduced fading and jamming. An interesting application here is the use of a noisy source, the noise output in the utilized spectrum from the source canceling the receiver noise output inherent in the transmission. Another application implies that the receiver may be provided with automatic tuning, synchronized to the transmitter, so that frequency drift may be minimized, and interception and jamming reduced by continuous periodic or random frequency shifts.

The above observations do not cover all phases or the possible uses of the reflected-power scheme, but indicate where the essential differences from conventional communication systems are to be sought.

The value of the reflected-power principle from an operational point of view has not yet been fully estimated, but the following applications may serve to illustrate the usefulness of the new method.

Meteorological balloon tracking and telemetering is conventionally done by means of a balloon transmitter and a directional search receiver. Alternatively, the balloon may be provided with a telemetering-data-modulated special reflector. A properly arranged radar on the

ground then serves the double purpose of tracking the balloon and receive the telemetering signal, so that the procedure of meteorological observation becomes simplified.

Microwave relaying is illustrated by Fig. 11. Fig. 11(a) shows a conventional communications relay link, and in 11(b) and 11(c) the same link provided with reflected power. In this case the unmodulated transmitter is located at B' and the modulated reflector at A'. The transmitter to the right has a sufficiently wide unmodulated beam, which always includes the modulated reflector to the left, but ideally the return radiation is always pinpointed on the receiver to the right, even in case of atmospheric bending. Thus reliable communication is maintained between A' and B', and the beam is still very sharp, assuming an ideal reflector.

The above arrangement can be extended to automatic beam stabilization by means of error-signal reflectors *a*, *b*, *c*, *d*, or by other similar arrangement.

Frequency-drift-free communication: In contrast to the conditions for conventional communication links, we can here apply afc to the transmitter, or both to the transmitter and the local oscillator. As the transmitter has an excess of power, simplified afc circuits become possible.

Fading-free communication can be established in various ways by means of the reflected-power communication scheme. It should be noticed that if a conventional communication link provides zero received signal because of interference, reflected power would provide maximum signal at the same instant for the same distance, as the path length is doubled. Combinations of the systems may thus eliminate fading.

An example of automatic fading elimination by means of frequency variation is shown in Fig. 12, where the transmitter and receiver are located to the left and the modulated reflector to the right. When the input signal to the receiver becomes small because of fading, an error signal is generated which, via a reactance tube or simply via rectification, operates on the transmitter, changing the frequency until proper transmission conditions are restored. Alternatively, a system with horizontally or vertically arranged error-signal reflectors may be used, such as the reflector system *a*, *b*, and the error signal made to operate the transmitter frequency until maximum signal strength in *m* is restored.

It is of interest to note that some of the uses of reflected power make possible the building of *receivers without local oscillator*. Sometimes, at microwaves, it is a great simplification to eliminate the local oscillator due to the difficulty of producing suitable oscillators. Sometimes, however, the noise increases so much when the local oscillator is removed that the conventional scheme is preferred.

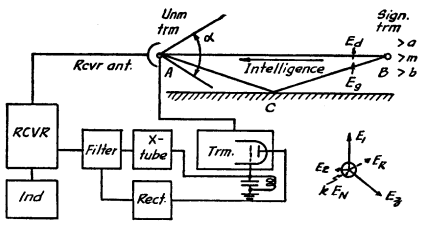

Fig. 12—An example of fading-free communication by means of the reflected power principle. Signal reduction due to interference is counteracted by an automatic change in frequency.

The simplest local-oscillator-free receiver circuit is the one where a part of the transmitter carrier is injected so that beats containing the signal are produced with the receiver input wave. An example of another arrangement for the elimination of the local oscillator is shown in Fig. 13, where an external oscillator and the transmitter feed into a nonlinear mixer. The frequency f_2 of the external oscillator is such that one of its harmonics kf_2 is suitable for use as the intermediate frequency. The nonlinear device output is f_1+kf_2, where f_1 is the transmitter frequency, and is injected into the receiver as local-oscillator frequency. By beating with the receiver input this local-oscillator wave provides the desired if output in the receiver nonlinear mixer.

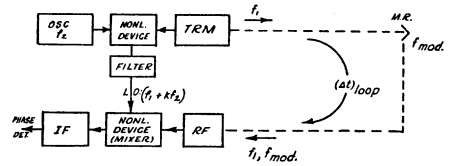

Fig. 13—A simple communications circuit showing the possibility of superheterodyne reception without the need of a local oscillator, the microwave local oscillator being replaced by an rf oscillator and associated mixers and filters.

Some of the schemes referred to above employ carrier injection in one form or another in the receiver input. This means that the coherence of the input signal can be enhanced and, accordingly, the signal-to-noise ratio improved.

Conclusion

It should be mentioned in conclusion that the reflected-power principle has already proven its value in the well-known radar application, i.e., for simple on-off modulation. In the method presented we modulate the target with any time function. Evidently considerable research and development work has to be done before the remaining basic problems in reflected-power communication are solved, and before the field of useful applications is explored.

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